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NEW PROVISIONAL APPLICATION TRANSMITTAL LETTER

Sir:

Transmitted herewith for filing is the Provisional Patent Application of Inventor(s):

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For: Equivalent Estimation Method For Evaluating Subscriber Lines Based on Time Domain Reflectometry

Enclosed are the following papers required to obtain a filing date under 37 C.F.R. §1.53(c):

__ Sheets of Informal Drawings
19 Pages of Specification, Drawings & Tables
Claims
8 Appendices

The following papers, if indicated by an ☒, are also enclosed:

☐ A Declaration and Power of Attorney
☐ An Assignment of the invention
☐ An Information-Disclosure Statement, Form PTO-1449 and a copy of each cited reference
☐ A Small-Entity Declaration
☒ A Certificate of Express Mailing, Express Mail Label No. ET 104264385 US

Basic Fee: $160

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UNITED STATES PROVISIONAL PATENT APPLICATION

Of

Ling Zheng and Michael Lund

for a

Equivalent Estimation Method for Evaluating Subscriber Lines Based On Time Domain Reflectometry

Equivalent Estimation Method for Evaluating Subscriber Lines based on Time domain Reflectometry

Ling Zheng, Michael Lund

I. Overview of an Exemplary Embodiment of the Invention

An exemplary embodiment of the present invention relates to a method for evaluating the channel capacity of a twisted-pair copper line and, more specifically, to a method for predicting the data rate of a subscriber line for potential digital subscriber line (DSL) service without explicitly estimating the topology of the line.

DSL technology makes it possible to transport high-bit-rate digital information via a subscriber line. The channel capacity, which is defined as the obtainable data rate from a given line, relates to the physical structure and topology of the line, such as the length of the line, gauge, existence of bridged taps, bridged tap locations and lengths, etc.. Thus, if the topology of the line is known, the data rate can be predicted prior to providing DSL service to the customer.

The time domain reflectometry (TDR) is a very useful tool for characterizing a subscriber line. It operates by sending an electrical pulse down to the line and measuring the returned signal, referred to as a TDR echo. The measured echo contains the information about the physical structure and topology of the line.

The most common method for evaluating the channel capacity from a TDR echo is by explicitly estimating the physical topology of the line based on the transmission line theory, and then searching a database to find the data rate corresponding to the specific topology. The problem with this method is the estimation complexity increases dramatically when the topology of the line is getting complicate. For example, consider a line with N consecutive sections of different gauges but without bridged taps. Because there are two variables needed to be estimated for each section, the gauge and the length, the searching space is 2^N dimensions. The computation complexity increases exponentially with the potential number of distinct sections.

As explained in the proceeding paragraphs, the ultimate goal of estimating the topology of a subscriber line from the TDR echo is to evaluate the channel capacity of the line. Hence, if one can estimate a line with a much simpler topology but the same channel capacity as the actual line from the TDR echo, the goal can be achieved with much less computation complexity. An exemplary embodiment of the present invention follows this idea, and provides a method of evaluating the channel capacity from the TDR echo without explicitly estimating the topology of the line.

For simplicity, we refer a subscriber line without bridged tap as a straight line. Here "straight" means no bridged tap presented. Also, a subscriber line is referred as a line, a loop or a wire. There is no difference among these terms.

II. Background of the invention

The present invention is developed based on the transmission line theory. A typical transmission line system can be schematically represented as in Fig.1, where V_s is the output voltage of the source, Z_s is the output impedance of the source, Z_L is the input impedance of the

load, and L is the length of the line connecting the source and the load. V_s, Z_s, and Z_L are functions of frequency. According to the transmission line theory, when the electrical wave generated by the source travels down to the line, it is attenuated by the line, and reflected whenever there is an impedance discontinuity. The wave travels forward and backward inside the line to infinity.

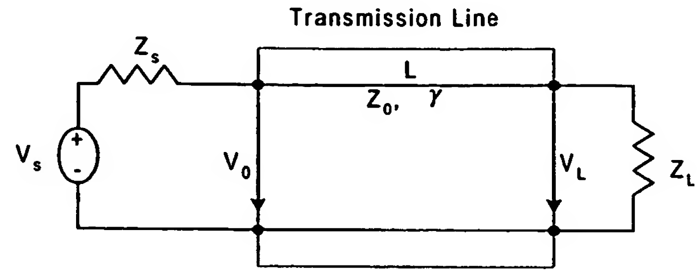

Figure 1: A schematic representation of a transmission line system with a single-section straight loop.

Assume the characteristic impedance and the propagation constant of a given line to be Z_0 and γ, where Z_0 and γ are functions of frequency. Consider a straight loop with a single gauge, the only impedance changes are (1) at the connection between the source and the line, and (2) at the connection between the line and the load. Assume the reflection coefficient, which is defined as the ratio of reflected to forward voltage, to be ρ_s at the source output, and to be ρ_L at the load, where

$$\rho_s = \frac{Z_s - Z_0}{Z_s + Z_0},$$

$$\rho_L = \frac{Z_L - Z_0}{Z_L + Z_0}.$$

According to the wave propagation theory, the voltage at the output of the source, denoted as V_0, can be represented by a combination of an incident wave and infinite number of multi-reflections caused by impedance discontinuity:

Incident wave: $$V_0^{(0)} = \frac{Z_0}{Z_0 + Z_s} V_s$$

1st reflection: Forward wave $V_0^{(1)+} = V_0^{(0)} \cdot e^{-2\gamma L} \cdot \rho_L \cdot \rho_s$,
Backward wave $V_0^{(1)-} = V_0^{(0)} \cdot e^{-2\gamma L} \cdot \rho_L$,

2nd reflection: Forward wave $V_0^{(2)+} = V_0^{(0)} \cdot e^{-4\gamma L} \cdot \rho_L^2 \cdot \rho_s^2$,
Backward wave $V_0^{(2)-} = V_0^{(0)} \cdot e^{-4\gamma L} \cdot \rho_L^2 \cdot \rho_s$,

3rd reflection: Forward wave $V_0^{(3)+} = V_0^{(0)} \cdot e^{-6\gamma L} \cdot \rho_L^3 \cdot \rho_s^3$,
Backward wave $V_0^{(3)-} = V_0^{(0)} \cdot e^{-6\gamma L} \cdot \rho_L^3 \cdot \rho_s^2$,

⋮ ⋮

n[th] reflection: Forward wave $V_0^{(n)+} = V_0^{(0)} \cdot e^{-2n\gamma L} \cdot \rho_L^n \cdot \rho_s^n$,

Backward wave $V_0^{(n)-} = V_0^{(0)} \cdot e^{-2n\gamma L} \cdot \rho_L^n \cdot \rho_s^{n-1}$.

The forward wave is the voltage wave traveling away from the source. The backward wave is the voltage wave traveling toward the source.

Therefore, V_0 can be expressed as

$$
\begin{aligned}
V_0 &= \sum_{n=0}^{\infty} V_0^{(n)} = V_0^{(0)} + \sum_{n=1}^{\infty} \left[V_0^{(n)+} + V_0^{(n)-} \right] \\
&= V_0^{(0)} + \sum_{n=1}^{\infty} V_0^{(0)} \cdot e^{-2n\gamma L} \cdot \rho_L^n \cdot \rho_s^{n-1} (1 + \rho_s) \\
&= V_s \cdot \frac{Z_0}{Z_0 + Z_s} \cdot \frac{1 + e^{-2\gamma L} \cdot \rho_L}{1 - e^{-2\gamma L} \cdot \rho_L \cdot \rho_s} .
\end{aligned}
\tag{1}
$$

Likewise, the voltage at the input of the load, denoted as V_L, can also be represented by a combination of multi-reflections:

1[st] reflection: Forward wave $V_L^{(1)+} = V_0^{(0)} \cdot e^{-\gamma L}$,

Backward wave $V_L^{(1)-} = V_0^{(0)} \cdot e^{-\gamma L} \cdot \rho_L$,

2[nd] reflection: Forward wave $V_L^{(2)+} = V_0^{(0)} \cdot e^{-3\gamma L} \cdot \rho_L \cdot \rho_s$,

Backward wave $V_L^{(2)-} = V_0^{(0)} \cdot e^{-3\gamma L} \cdot \rho_L^2 \cdot \rho_s$,

3[rd] reflection: Forward wave $V_L^{(3)+} = V_0^{(0)} \cdot e^{-5\gamma L} \cdot \rho_L^2 \cdot \rho_s^2$,

Backward wave $V_L^{(3)-} = V_0^{(0)} \cdot e^{-5\gamma L} \cdot \rho_L^3 \cdot \rho_s^2$,

⋮ ⋮

n[th] reflection: Forward wave $V_L^{(n)+} = V_0^{(0)} \cdot e^{-2(n-1)\gamma L} \cdot \rho_L^{n-1} \cdot \rho_s^{n-1}$,

Backward wave $V_L^{(n)-} = V_0^{(0)} \cdot e^{-2(n-1)\gamma L} \cdot \rho_L^n \cdot \rho_s^{n-1}$.

Therefore, V_L can be expressed as

$$
\begin{aligned}
V_L &= \sum_{n=1}^{\infty} V_L^{(n)} = \sum_{n=1}^{\infty} \left[V_L^{(n)+} + V_L^{(n)-} \right] \\
&= \sum_{n=1}^{\infty} V_0^{(0)} \cdot e^{-(2n-1)\gamma L} \cdot \rho_L^{n-1} \cdot (1+\rho_L) \cdot \rho_s^{n-1} \\
&= V_s \cdot \frac{Z_0}{Z_0 + Z_s} \cdot \frac{(1+\rho_L) \cdot e^{-\gamma L}}{1 - e^{-2\gamma L} \cdot \rho_L \cdot \rho_s} .
\end{aligned} \tag{2}
$$

Both a TDR measurement system and a DSL application can be represented by the equivalent circuitry shown in Fig.1. In the TDR case, V_s is the pulse sent down to the line, V_0 is the measured TDR echo. The incident wave in V_0 is referred to as the near-end echo, the sum of the multi-reflections is referred to as the far-end echo. In a TDR measurement, the source impedance is usually the same as the characteristic impedance of the line, i.e. $Z_s = Z_0$, and the end of the line is usually open, i.e. $Z_L = \infty$, thus $\rho_s = 0$, $\rho_L = 1$, and the measured TDR echo can be written as

$$
V_0 = V_s \cdot \frac{Z_0}{Z_0 + Z_s} \cdot \left(1 + e^{-2\gamma L}\right), \tag{3}
$$

which is a combination of the near-end echo and the backward wave in the 1st reflection.

In the DSL application case, V_s and Z_s represent the equivalent circuitry of the modem at the central office (CO), and Z_L represents the equivalent circuitry of the modem at the customer premise (CPE). The obtainable data rate relates to the transfer function of the subscriber line, which is defined as

$$
H = \frac{V_L}{V_0} .
$$

A modem is usually designed to have an impedance matching to the line, i.e. $Z_s = Z_0$, $Z_L = Z_0$, thus $\rho_s = 0$, $\rho_L = 0$, and V_0, V_L, and H can be written as

$$
\begin{aligned}
V_0 &= V_s \cdot \frac{Z_0}{Z_0 + Z_s}, \\
V_L &= V_s \cdot \frac{Z_0}{Z_0 + Z_s} \cdot e^{-\gamma L}, \\
H &= e^{-\gamma L} .
\end{aligned} \tag{4}
$$

Because the imaginary part of γ is a linear function of frequency, H has linear phase, the data rate is mainly determined by the modular of the transfer function.

Assume the transfer functions of two single-gauge straight lines to be H_1 and H_2. According to Eq. (4),

$$
\begin{aligned}
H_1 &= e^{-\gamma_1 L_1}, \\
H_2 &= e^{-\gamma_2 L_2},
\end{aligned}
$$

where γ_1 and L_1 are the propagation constant and length of Line 1, and γ_2 and L_2 are the propagation constant and length of Line 2. These two lines will have the same data rate, say, to be equivalent, if

$$
real(\gamma_1) \cdot L_1 = real(\gamma_2) \cdot L_2 , \tag{5}
$$

where $real(\cdot)$ is an operation to get a variable's real part.

Assume the TDR echoes of these two lines to be V_{01} and V_{02}. According to Eq. (3),

$$V_{01} = V_s \cdot \frac{Z_{01}}{Z_{01} + Z_s} \cdot \left(1 + e^{-2\gamma_1 L_1}\right),$$

$$V_{02} = V_s \cdot \frac{Z_{02}}{Z_{02} + Z_s} \cdot \left(1 + e^{-2\gamma_2 L_2}\right),$$

where Z_{01} and Z_{02} are the characteristic impedance of Line 1 and Line 2. If one can ignore the difference in characteristic impedance between these two lines, i.e. $Z_{01} \approx Z_{02} = Z_0$, these two lines have similar near-end echoes, then according to Eq. (5), when these two lines have the same data rate, the following equation holds true:

$$\left|V_s \frac{Z_0}{Z_0 + Z_s}\right| \cdot e^{-2 \cdot real(\gamma_1) L_1} = \left|V_s \frac{Z_0}{Z_0 + Z_s}\right| \cdot e^{-2 \cdot real(\gamma_2) L_2} . \tag{6}$$

The left-hand side of Eq. (6) is the amplitude of the far-end TDR echo of Line 1; the right hand side is that of Line 2. Eq. (6) indicates that the far-end TDR echoes from two single-gauge straight lines, which have the same data rate, have the same amplitude. Although this deduction is derived for certain Z_s and Z_L, it holds true in general. The present invention is developed based on this deduction.

An exemplary embodiment of the present invention focuses on estimating data rates for asymmetric DSL (ADSL) service. ADSL has an upstream (US) band, within which the data is transmitted from the CPE to CO, from Tone 6 to Tone 31, and a downstream (DS) band, within which the data is transmitted from the CO to CPE, from Tone 32 to Tone 255. The tone interval is 4312.5Hz, say, the ith tone corresponds to frequency $f_i = i \times 4312.5$ (Hz). The TDR echo is measured by an ADSL CO modem within one frame, and is averaged over 10,000 frames. Each frame has 512 time samples. The sampling rate is 2208kHz. All subscriber lines are equivalent to a single-gauge 26AWG straight loop.

III. Description of an exemplary embodiment of the Invention

3.1 Straight loops with a single gauge

For a straight loop with a single gauge, Gauge x, let the propagation constant be γ_x, the physical length be L_x, then the equivalent equation given in Eq. (5) can be rewritten as

$$real(\gamma_x) \cdot L_x = real(\gamma_{26}) \cdot L_{26} , \tag{7}$$

where γ_{26} and L_{26} is the propagation constant and loop length of the corresponding 26AWG equivalent loop. Let the equivalent loop length ratio between Gauge x and 26AWG be a_x, say, $L_x = a_x \cdot L_{26}$, then Eq. (7) becomes

$$real(\gamma_x) \cdot a_x = real(\gamma_{26}) . \tag{8}$$

Because the propagation constant varies across frequency, a fixed ratio across frequency has to be computed in a least square sense over a certain frequency band:

$$a_x(m,n) = \frac{\sum_{i=m}^{n} real[\gamma_x(f_i)] \cdot real[\gamma_{26x}(f_i)]}{\sum_{i=m}^{n} \{real[\gamma_x(f_i)]\}^2} , \tag{9}$$

where f_i is the frequency of the ith tone, and m and n determine the frequency band under consideration. The equivalent ratio changes with m and n.

Because a subscriber line provides more attenuation on high frequency components than on low frequency components, the measured TDR echo is dominated by low frequency components. The spectrum of the pulse also influences the frequency band of the measured TDR echo. Fig. 2 shows the spectrum of the TDR pulse used in this invention. Fig. 3 shows the spectra of the far-end echoes from 26AWG straight loops with various lengths. It can be seen that the far-end echo is dominated by low frequency components, and more than 95% of the energy is distributed below Tone 50 (≅220kHz). Since there is no much energy below Tone 6, the TDR band is set from Tone 6 to 50.

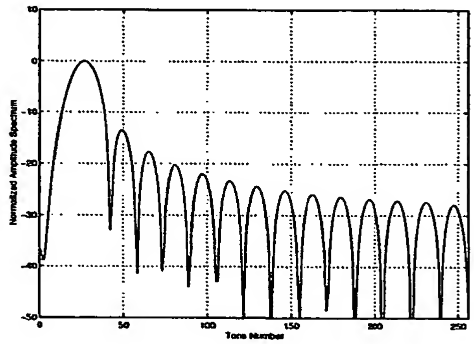

Figure 2: Spectrum of the TDR pulse used in this invention.

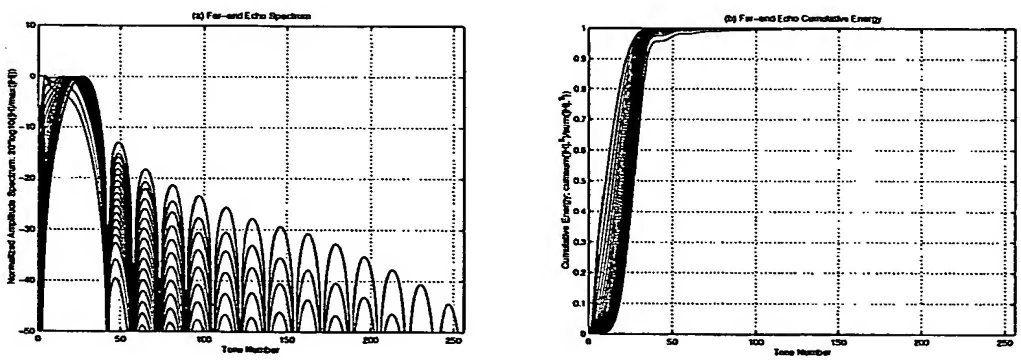

Figure 3: Far-end echoes of 26AWG straight loops in the frequency domain. Eighteen curves are plotted in each plot. Each curve corresponds to a 26AWG loop with a certain length. The loop length varies from 1kft to 18kft in 1kft step.

As mentioned above, the US band of an ADSL application is from Tone 6 to Tone 31, the DS band is from Tone 32 to Tone 255, so the equivalent ratio for the US case is different from the DS case. Consider the difference in frequency band among the US, DS, and TDR case, define

$a_{x(eq\_us)}$ = equivalent loop length ratio corresponding to the same US data rate,

$a_{x(eq\_ds)}$ = equivalent loop length ratio corresponding to the same DS data rate,

$a_{x(eq\_tdr)}$ = equivalent loop length ratio corresponding to the same far-end echo (shape and amplitude only)

then the equation for computing each ratio using Eq. (9) is

$$a_{x(eq\_us)} = a_x(m=6,\ n=31)\ ,$$
$$a_{x(eq\_ds)} = a_x(m=32,\ n=255)\ ,$$
$$a_{x(eq\_tdr)} = a_x(m=6,\ n=50)\ .$$

Table 1 shows the equivalent ratios for both American loops (AWG) and European loops (metric). All ratios are computed using the wire primary parameters, characteristic impedance Z_0 and propagation constant γ, published in ITU G. 996.1.

Table 1: Equivalent coefficients

Gauge x	Equivalent-US-Rate Ratio $a_{x(eq\_us)}$	Equivalent-DS-rate Ratio $a_{x(eq\_ds)}$	Equivalent-TDR-echo Ratio $a_{x(eq\_tdr)}$	Velocity Coefficient V_x *(sample/kft)*	Time-shift Coefficient τ_x *(ft/sample)*
0.32mm	0.8	0.8	0.8	6.3	ANY
0.4mm	1.0	1.0	1.0	6.9	ANY
0.5mm	1.5	1.3	1.5	7.5	30
0.63mm	2.2	1.6	2.0	6.9	40
0.9mm	3.3	2.2	3.0	6.5	30
19AWG	3.2	2.3	2.8	6.9	20
22AWG	2.0	1.6	1.9	7.1	30
24AWG	1.4	1.27	1.4	7.1	40
26AWG	1.0	1.0	1.0	7.0	ANY

Fig.4 shows the data rate comparison between 24AWG and 26AWG equivalent-data-rate loops. According to Table 1, the equivalent-US-rate ratio is 1.4; the equivalent-DS-rate ratio is 1.27. Fig. 4(a) shows the US data rates of 24AWG loops (length = *L*) versus that of 26AWG equivalent-US-rate loops (length = L/1.4). Fig. 4(b) shows the DS data rates of 24AWG loops (length = L) versus that of 26AWG equivalent-DS-rate loops (length = L/1.27). It can be seen that the equivalent loops do have very similar data rate. The averaged data rate difference between equivalent loops is about 30 kbps for the US case, and 160 kbps for the DS case.

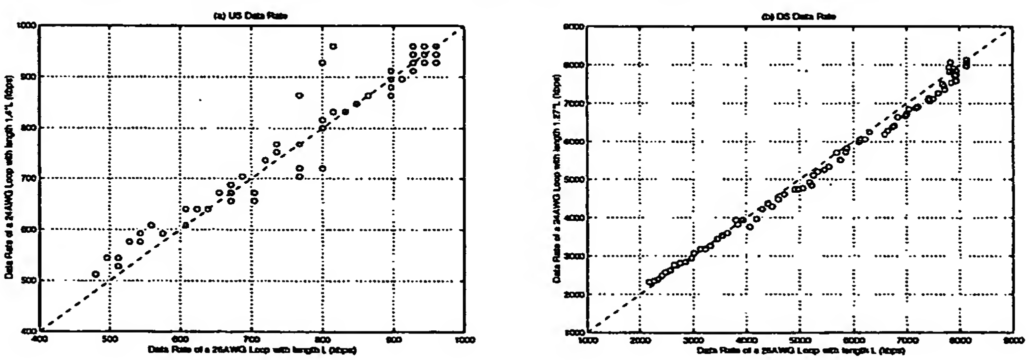

Figure 4: Data rate comparison between equivalent-data-rate loops. (a) US data rate. (b) DS data rate.

Fig. 5 shows the far-end echo comparison between 24AWG and 26AWG equivalent-TDR-echo loops. According to Table 1, the equivalent-TDR-echo ratio is 1.4. Fig. 5 (a) shows the comparison between a 26AWG 5kft loop and a 24AWG 7kft loop (5*1.4=7). Fig. 5 (b) shows the comparison between a 26AWG 10kft loop and a 24AWG 14kft loop (10*1.4=14). It can be seen that between the equivalent loops, the amplitude and the shape of the far-end echoes are very similar; while the time delays are different - the 24AWG loop has a longer delay than the 26AWG loop.

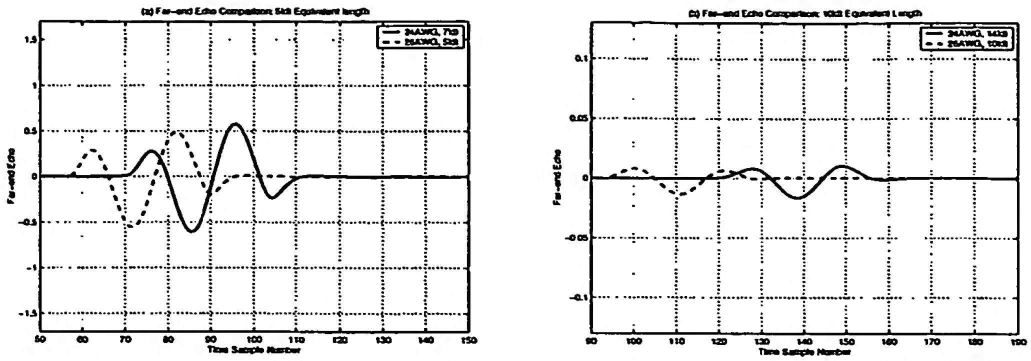

Figure 5: Far-end echo comparison between 24AWG and 26AWG equivalent-TDR-echo loops. (a) 26AWG equivalent loop length = 5kft. (b) 26AWG equivalent loop length = 10kft.

The equivalent ratios given in Table 1 are computed under the assumption that the difference in characteristic impedance Z_0 can be ignored. The comparisons shown in Fig. 4 and Fig. 5 indicate this approximation is reasonable.

3.2 Straight loops with mixed gauges

Because of the existence of gauge changes within a mixed-gauge loop, the mixed-gauge case is more complicate than the single gauge case. A typical transmission line system with a two-section mixed-gauge loop is schematically represented as in Fig. 6. It is mostly the same as that shown in Fig. 1. The only difference is the line has two sections instead of one section in this plot. Let the characteristic impedance, propagation constant, and loop length of the first section be Z_{01}, γ_1 and L_1, and that of the second section be Z_{02}, γ_2 and L_2. Denote the gauge change point as A, and the reflection coefficient at A as ρ_A, where

$$\rho_A = \frac{Z_{02} - Z_{01}}{Z_{02} + Z_{01}}.$$

According to the transmission line theory, the wave propagation within this two-section loop can be represented as in the following:

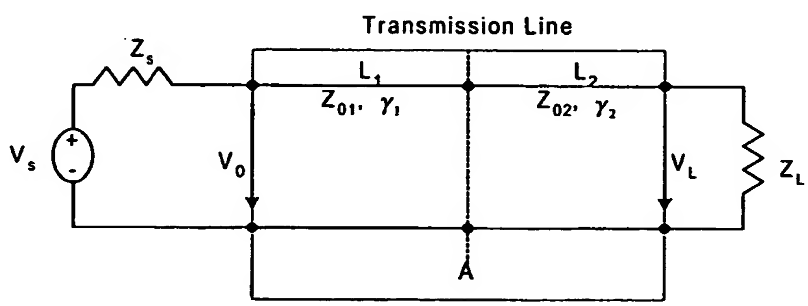

Figure 6: A schematic representation of a transmission line system with a 2-section straight loop.

Incident wave:

$$V_0^{(0)} = \frac{Z_{01}}{Z_s + Z_{01}} V_s.$$

1st trip, wave at A:

Forward wave $V_{A12}^{(1)+} = V_0^{(0)} \cdot e^{-\gamma_1 L_1}$

Backward wave $V_{A12}^{(1)-} = V_0^{(0)} \cdot e^{-\gamma_1 L_1} \cdot \rho_A$

1st trip, incident from Section 1 to Section 2:

$$V_{A12}^{(1)} = V_{A12}^{(1)+} + V_{A12}^{(1)-} = V_0^{(0)} \cdot e^{-\gamma_1 L_1} \cdot (1+\rho_A)$$

1st trip, return from the end of Section 2 to A:

Forward wave $V_{A21}^{(1)+} = V_{A12}^{(1)} \cdot e^{-2\gamma_2 L_2} \cdot (-\rho_A) \cdot \rho_L$

Backward wave $V_{A21}^{(1)-} = V_{A12}^{(1)} \cdot e^{-2\gamma_2 L_2} \cdot \rho_L$

1st trip, incident from Section 2 to Section 1:

$$V_{A21}^{(1)} = V_{A21}^{(1)+} + V_{A21}^{(1)-}$$

1st reflection at the source output:

$$\begin{aligned} V_0^{(1)} &= \left[V_{A12}^{(1)-} \cdot e^{-\gamma_1 L_1} + V_{A21}^{(1)} \cdot e^{-\gamma_1 L_1}\right] \cdot (1+\rho_s) \\ &= \left[V_{A12}^{(1)-} \cdot e^{-\gamma_1 L_1} + \left(V_{A21}^{(1)+} + V_{A21}^{(1)-}\right) \cdot e^{-\gamma_1 L_1}\right] \cdot (1+\rho_s) \\ &= \left[V_0^{(0)} \cdot e^{-2\gamma_1 L_1} \cdot \rho_A + V_0^{(0)} \cdot e^{-2\gamma_1 L_1 - 2\gamma_2 L_2} \cdot (1-\rho_A^2) \cdot \rho_L\right] \cdot (1+\rho_s) \\ &= V_s \frac{Z_{01}}{Z_{01}+Z_s} \cdot e^{-2\gamma_1 L_1} \cdot \rho_A \cdot (1+\rho_s) \\ &\quad + V_s \frac{Z_{01}}{Z_{01}+Z_s} \cdot e^{-2\gamma_1 L_1 - 2\gamma_2 L_2} \cdot (1-\rho_A^2) \cdot \rho_L \cdot (1+\rho_s) \end{aligned} \tag{10}$$

When each section is not too short (L_1, $L_2 \geq$ 1000ft), higher order reflections can be ignored, thus the 1st order reflection given by Eq. (10) is a reasonable approximation of the overall far-end echo. Eq, 10 shows that the far-end echo includes two dominant reflections, one is from the gauge change (the first term in Eq. (10)), the other is from the end of the line (the second term in Eq. (10)). Because the reflection coefficient at the gauge change, ρ_A, is usually very small, ρ_A^2 is even smaller, the reflection from the end of the line can be simplified as

$$V_s \cdot \frac{Z_{01}}{Z_{01}+Z_s} \cdot e^{-2\gamma_1 L_1 - 2\gamma_2 L_2} \cdot \rho_L \cdot (1+\rho_s). \tag{11}$$

Assume the equivalent-TDR ratio to a 26AWG loop to be $a_{1(eq\_tdr)}$ for Section 1, and $a_{2(eq\_tdr)}$ for Section 2. Eq. (11) indicates that if Z_{01} is similar to the characteristic impedance of a 26AWG loop, the reflection from the end of the mixed-gauge loop has similar shape and amplitude as the far-end echo from a 26AWG loop with length

$$L_{eq\_tdr} = L_1 / a_{1(eq\_tdr)} + L_2 / a_{2(eq\_tdr)}.$$

Fig. 7 shows far-end echo comparisons between a mixed-gauge loop and its 26AWG equivalent-TDR-echo loop. Fig. 7 (a) shows the comparison between a mixed-gauge loop, which has a first section of 3kft 26AWG and a second section of 3kft 24AWG, and a 26AWG straight loop with a length of 5.1kft (3+3/1.4=5.1). Fig. 7 (b) shows the comparison between a mixed-gauge loop, which has a first section of 2kft 24AWG and a second section of 5kft 26AWG, and a 26AWG straight loop with a length of 6.4kft (2/1.4+5=6.4). Fig. 7(a) suggests that in the mixed-gauge case, although part of the far-end echo is the return from the gauge change, the return from

the end of the loop is very similar to the far-end echo from the equivalent straight loop. It is worth to mention that the sign of the return from the gauge change is the inverse of the return from the end of the loop. This is consistent with the fact that the gauge change is from 26AWG to 24AWG. Because the characteristic impedance of a 24AWG loop is smaller than that of a 26AWG loop, the reflection coefficient at the gauge change is negative, which results in a negative return. Similar to Fig. 7(a), Fig. 7 (b) also suggests that in the mixed-gauge case, although part of the far-end echo is the return from the gauge change, the return from the end of the loop is very similar to the far-end echo from the equivalent straight loop. However, now the gauge change is from 24AWG to 26AWG, the reflection coefficient at the gauge change is positive, so the return from the gauge change is somewhat similar to the return from the end of the loop. Therefore, it is expected that the return from the gauge change may have more influence on 24AWG+26AWG mixed-gauge case than on 26AWG+24AWG mixed-gauge case.

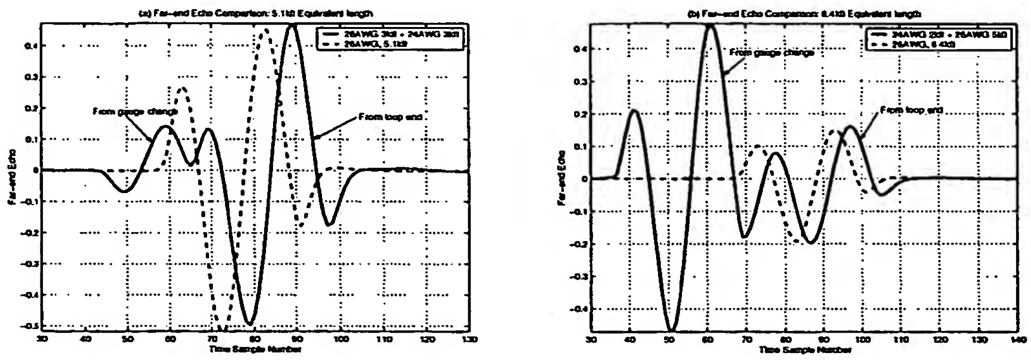

Fig. 7 Far-end echo comparison between mixed-gauge loops and 26AWG equivalent loops. (a) The mixed-gauge loop = 26AWG 3kft + 24AWG 3kft. The 26AWG straight loop = 5.1kft. (b) The mixed-gauge loop = 24AWG 2kft + 26AWG 5kft. The 26AWG straight loop = 6.4 kft.

Assume $Z_s = Z_{01}$, $Z_L = Z_{02}$, the voltage at the load input, V_L, is

$$V_L = V_{A12}^{(1)} \cdot e^{-\gamma_2 L_2} = V_0^{(0)} \cdot e^{-\gamma_1 L_1 - \gamma_2 L_2} \cdot (1+\rho_A),$$

therefore, the transfer function of the mixed-gauge line is

$$H = \frac{V_L}{V_0} = e^{-\gamma_1 L_1 - \gamma_2 L_2} \cdot (1+\rho_A). \tag{12}$$

As mentioned before, the reflection coefficient at the gauge change, ρ_A, is usually very small, the transfer function can be simplified as

$$H \approx e^{-\gamma_1 L_1 - \gamma_2 L_2}. \tag{13}$$

Assume the equivalent US and DS ratios to a 26AWG loops to be $a_{1(eq\_us)}$ and $a_{1(eq\_ds)}$ for Section 1, and $a_{2(eq\_us)}$ and $a_{2(eq\_ds)}$ for Section 2. Eq. 13 indicates the 2-section loop has the same data rate as a 26AWG straight loop with length

$$L_{eq\_us} = L_1 / a_{1(eq\_us)} + L_2 / a_{2(eq\_us)},$$

and has the same DS data rate as a 26AWG straight loop with length

$$L_{eq\_ds} = L_1 / a_{1(eq\_ds)} + L_2 / a_{2(eq\_ds)}.$$

Fig. 8 shows data rate comparison between 2-section mixed-gauge loops and their 26AWG equivalent-data-rate loops. The first section of each loop is 26AWG, the second section is 24AWG. Each section varies from 1kft to 9kft in 500ft step. It can be seen that the approximation given in Eq. (13) is reasonable.

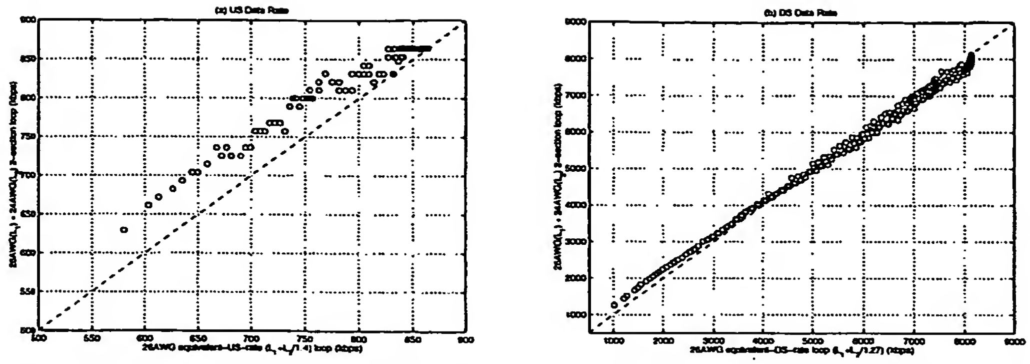

Figure 8: Data rate comparison between 2-section mixed-gauge loops and their 26AWG equivalent loops. (a) US data rate. (b) DS data rate.

Fig. 9 shows the data rate comparison between 2-section loops with different gauge change order: one case is from 26AWG to 24AWG; the other case is from 24AWG to 26awg. It can be seen that the data rates of these two cases are very similar. According to Eq. (12), the transfer function of the 26AWG+24AWG case be written as

$$H_{26AWG+24AWG} = e^{-\gamma_{26}L_{26}-\gamma_{24}L_{24}} \cdot (1+\rho_A),$$

and the 24AWG+26AWG case be written as

$$H_{24AWG+26AWG} = e^{-\gamma_{24}L_{24}-\gamma_{26}L_{26}} \cdot (1-\rho_A),$$

thus the difference between these two cases is the sign of the reflection coefficient. The similarity in data rates between these two cases indicates the reflection at the gauge change has no significant influence on the data rates. Therefore, Eq. (13) is a reasonable approximation of the transfer function of a 2-section loop.

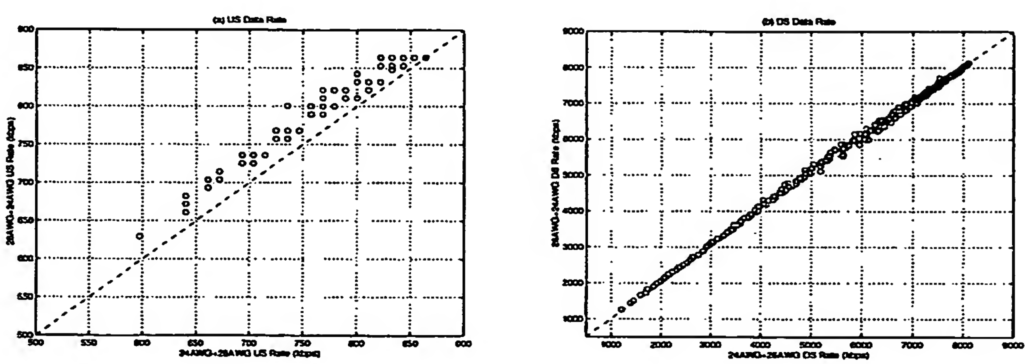

Figure 9: Data rate comparison between 26AWG+24AWG loops and 24AWG+26AWG loops. (a) US data rate. (b) DS date rate.

Although Eq. (11) and Eq. (13) are derived for 2-section loops, they hold true for multi-section loops. For a loop with *n* sections, assume the physical length, equivalent US ratio, equivalent DS ratio, and equivalent TDR ratio for the *ith* section ($i = 1$ - n) to be L_i, $a_{i(eq\_us)}$, $a_{i(eq\_ds)}$, and $a_{i(eq\_tdr)}$, respectively. Let the 26AWG equivalent US length, equivalent DS length, and equivalent TDR length be L_{eq_us}, L_{eq_ds}, and L_{eq_tdr}, respectively, then

$$L_{eq\_us} = \sum_{i=1}^{n} L_i / a_{i(eq\_us)},$$

$$L_{eq\_ds} = \sum_{i=1}^{n} L_i / a_{i(eq\_ds)},$$

$$L_{eq\_tdr} = \sum_{i=1}^{n} L_i / a_{i(eq\_tdr)}.$$

3.3 Correction of the equivalent length

As mentioned above, an exemplary goal of the equivalent loop estimation is to predict the data rate for a given loop. The equivalent length estimated from the measured TDR echo is the equivalent TDR length. According to Table 1, the equivalent TDR ratio and the equivalent data rate ratio, especially in the DS case, are different, thus the data rate predicted using the equivalent TDR length would be inaccurate. In accordance with an exemplary embodiment of the present invention, the equivalent-data-rate length is derived from the estimated equivalent TDR length and the time delay between the measured far-end echo and the equivalent far-end echo.

The time delay of a far-end echo is determined by the physical length of the loop and the propagation velocity of the traveling wave. Based on the transmission line theory, the time delay can be represented as

$$Delay = \frac{2L}{V_p},$$

where L is the physical length of the loop, and V_p is the propagation velocity. Since a far-end echo is a round trip return, the numerator is double loop length. The propagation velocity of a transmission line relates to the imaginary part of the propagation constant γ,

$$V_p = \frac{\omega}{imag(\gamma)},$$

where ω is the radian frequency $(\omega = 2\pi f)$, and $imag(\gamma)$ is the imaginary part of γ. Because a subscriber line usually has a propagation constant whose imaginary part is a straight line across frequency, the propagation velocity is a constant across frequency. Thus, the propagation velocity of Gauge x, denoted as V_x, can be computed from the propagation constant as in the following

$$V_x = \frac{2\pi \cdot \Delta f}{\frac{1}{n-m}\{imag[\gamma_x(f_n)] - imag[\gamma_x(f_m)]\}}, \qquad (14)$$

where f_i is the frequency of the *ith* tone, γ_x is the propagation constant of Gauge x, Δf is the tone interval, which in the ADSL case is 4312.5Hz, and m and n correspond to the frequency range of the TDR echo. In this specific case, $m = 6$, $n = 50$. The unit for V_x is "m/s".

The propagation velocity can be expressed in terms of number of samples per kft. For Gauge x, define number of samples per kft as velocity coefficient λ_x, then

$$\lambda_x = \frac{2 \times 1000(ft)}{3.2808(ft/m)} \cdot \frac{f_s}{V_x},$$

where f_s is the sampling rate of the TDR measurement system. The unit for λ_x is "Time Sample/kft". The velocity coefficient of AWG loops and metric loops are listed in Table 1 for f_s = 2208kHz. It can be seen that a 24AWG loop has almost the same velocity coefficient as a 26AWG loop, about 7 samples/kft. Consider the equivalent TDR ratio between a 26AWG loop and a 24AWG loop is 1.4, the delay difference, or the time shift, between a 24AWG loop (length

$= L$) and its 26AWG equivalent loop (length = $L/1.4$) is L×7 - L/1.4×7 = 2×L, where L is in kft. For the far-end echoes shown in Fig. 5 and Fig. 7, the theoretical time shifts between the measured far-end echo and its TDR-equivalent loop are:

Fig. 5(a): 24AWG length = 7 kft, time shift = 14 samples;
Fig. 5(b): 24AWG length = 10 kft, time shift = 28 samples;
Fig. 7(a): 26AWG length = 3 kft, 24AWG length = 3 kft, time shift = 6 samples;
Fig. 7(b): 24AWG length = 2 kft, 26AWG length = 5 kft, time shift = 4 samples.

The time shifts given above are consistent with the plots given in Fig. 5 and 7. This indicates the time shift contains the information about the physical length of a given loop.

For a straight loop with a single gauge, Gauge x, if its physical length is L_x, then the equivalent-TDR length $L_{eq\_tdr}= L_x/a_{x(eq\_tdr)}$. Let the velocity coefficient of Gauge x be λ_x, and that of 26AWG be λ_0, then the time shift between the measured echo and the equivalent echo, denoted as s, is

$$s = L_x \cdot \lambda_x - L_{eq\_tdr} \cdot \lambda_0,$$

where the unit of s is number of samples. The physical length, L_x, can be rewritten as

$$L_x = \frac{s}{\lambda_x - \lambda_0 / a_{x(eq\_tdr)}}. \tag{15}$$

Eq. (15) shows that once Gauge x is known, the physical length of a straight loop, L_x, can be obtained directly from the time shift s.

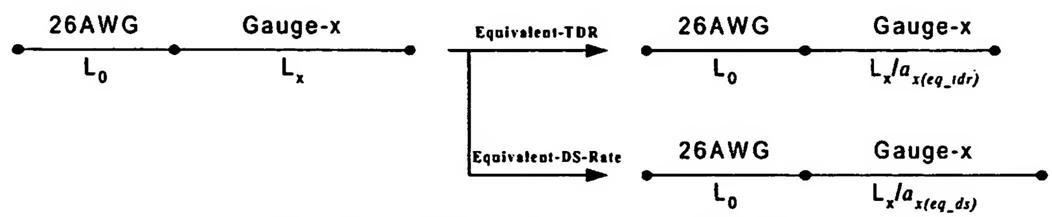

Figure10: A straight loop with mixed gauges.

For a straight loop mixed with 26AWG and Gauge x, as illustrated in Fig. 10, assume the length of the 26AWG section to be L_0 and that of Gauge x to be L_x. If we know λ_0, λ_x, $a_{x(eq\_tdr)}$, $a_{x(eq\_ds)}$, the equivalent-TDR length L_{eq_tdr}, and the time shift s, then the equivalent-DS-rate length, denoted as L_{eq_ds}, can be obtained by solving the following equations:

$$\begin{cases} L_{eq\_tdr} = L_0 + L_x / a_{x(eq\_tdr)} \\ L_{eq\_ds} = L_0 + L_x / a_{x(eq\_ds)} \\ L_x = s / \left(\lambda_x - \lambda_0 / a_{x(eq\_tdr)}\right) \end{cases}.$$

The solution is

$$L_{eq\_ds} = L_{eq\_tdr} + s \cdot \frac{1}{\lambda_x - \lambda_0 / a_{x(eq\_tdr)}} \cdot \left[\frac{1}{a_{x(eq\_ds)}} - \frac{1}{a_{x(eq\_tdr)}}\right]. \tag{16}$$

Define the time-shift coefficient for Gauge x be τ_x, i.e.

$$\tau_x = \frac{1}{\lambda_x - \lambda_0 / a_{x(eq\_tdr)}} \cdot \left[\frac{1}{a_{x(eq\_ds)}} - \frac{1}{a_{x(eq\_tdr)}}\right],$$

then Eq. (16) can be written as

$$L_{eq\_ds} = L_{eq\_tdr} + s \cdot \tau_x . \tag{17}$$

Eq. (17) shows the equivalent-data-rate length, L_{eq_ds}, can be determined based on the equivalent-TDR-echo length, L_{eq_tdr}, the time shift s, and the time shift coefficient τ_x. Table 1 lists the time shift coefficients of both AWG wires and metric wires. Because for 0.4mm and 26AWG wires, the time shift is 0 for any loop length, in other words, the equivalent-data length, the equivalent-TDR-echo length and the physical length are the same for these wires, the time-shift coefficient for these two gauges can be any value. Eq. (17) indicates that if the time shift coefficient, τ_x, is a constant across all gauges, the relationship given by Eq. (17) would be independent of Gauge x. However, Table 1 shows the time-shift coefficients are not identical across gauge. Since the most popular gauges used in the field are 24AWG and 26AWG in North American, and 0.4mm, 0.5mm, and 0.63mm in Europe, we average the time shift coefficient across 24AWG, 0.5mm and 0.63mm, the rounded average is $\tau_{mean} \approx 40$.

Because the equivalent-US-rate ratio is very similar to the equivalent-TDR ratio, no correction is made for US rate prediction.

IV. Description of the estimation procedure

An exemplary goal of the equivalent estimation method is to predict the data rate for a given line, such as a subscriber line, based on a TDR measurement. The input of the method is a measured TDR echo, the output is the predicted DS and US data rates. The intermediate steps include the equivalent TDR length estimation and the length correction for data rate prediction. In order to predict the data rate correctly, this method needs to know the data rate versus loop length curve of 26AWG straight loops. The exemplary detailed procedure is given step-in-step in the following.

Step1: Read in the measured TDR echo, denote it as $echo_{measure}(i)$, where i is the time sample index. In the ADSL case, one frame has 512 samples, $i = 0 - 511$.

Step 2: Calculate theoretical TDR echoes for 26AWG straight loops with various loop lengths.

Let the loop length of the nth loop be L_n, $n = 1 - N$, where N is total number of loops, the corresponding theoretical TDR echo be $echo_{model}(n, i)$, then

$$echo_{model}(n,i) = IFFT\left[\frac{Z_0 / \tanh(\gamma L_n)}{Z_s + Z_0 / \tanh(\gamma L_n)} V_s\right],$$

where Z_0 and γ are the characteristic impedance and propagation constant of a 26AWG loop, Z_s and V_s are the output impedance and voltage of the source.

Step 3: Isolate the theoretical far-end echo from the theoretical TDR echo and save the part of the far-end echo that includes most of the far-end energy.

The near-end echo of a 26AWG loop, denoted as $echo_{model\_near}(i)$, is

$$echo_{model\_near}(i) = IFFT\left[\frac{Z_0}{Z_s + Z_0} V_s\right].$$

The far-end echo, denoted as $echo_{model\_far}(n, i)$, is the difference between the TDR echo and the near-end echo,

$$echo_{model\_far}(n,i) = echo_{model}(n,i) - echo_{model\_near}(i).$$

As illustrated in Fig. 11, a far-end echo only lasts a short period of time, therefore, a time window is applied to each far-end echo, and only the waveform within this window is saved for later use. The start and stop points of this window vary with the loop length and shape of the input pulse. We define the time sample at which the cumulative energy exceeds 2% of the overall far-end echo energy as the start of the window, and fix the length of the window for all loops. Usually, the waveform within the time window contains more than 95% of the overall far-end echo energy. Let the start window of the *nth* loop be *st(n)*, the window size be *WN*, the windowed far-end echo be $y_n(j)$, then

$$y_n(j) = echo_{model\_far}(n, st(n)+j), \qquad (j = 0 - WN-1).$$

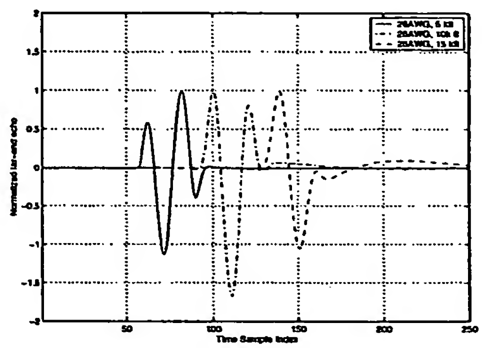

Figure 11: Far-end echoes from a few 26AWG straight loops.

Step 4: Estimate the equivalent TDR length.

The estimation procedures can be further divided into 5 steps:

(1) Isolate the far-end echo from the measured TDR echo, denote the isolated far-end echo as $x(i)$, $i = 0\text{-}511$. This isolation can be done by using several different methods:

a) Design an input pulse so that the far-end echo and the near-end echo are well separated in the time domain. In other words, the near-end echo has very low energy in the far-end echo window.

b) Remove the near-end echo using the difference between the far-end echo and the near-end echo, such as smoothness or spectrum distribution, etc..

(2) Let n = 1.

(3) Find the best time shift between the measured far-end echo and the theoretical far-end echo of the *nth* loop by solving the following optimization problem:

$$\min_m \sum_{j=0}^{WN-1} [x(j+m) - y_n(j)]^2 ,$$

where *m* is a variable representing the time shift, which varies in a certain region with a one-sample step. Denote the minimum error across *m* as *E(n)*, and the corresponding best shift as *S(n)*.

(4) Let $n = n+1$. If $n \le N$, go to (3); otherwise go to (5).

(5) Let $n^* = \min_n E(n)$, then the equivalent TDR length, L_{eq_tdr}, is

$$L_{eq\_tdr} = L_{n^*},$$

the corresponding time shift, denoted as s^*, is

$$s^* = S(n^*).$$

Step 5: Correct the equivalent-TDR length for data rate prediction.

According to the analysis given above, the equivalent-US-rate length and the equivalent-DS-rate length relate to L_{eq_tdr} and s^*:

$$L_{eq\_us} = L_{eq\_tdr}\text{, and}$$

$$L_{eq\_ds} = L_{eq\_tdr} + s^* \cdot \tau_{mean} = L_{eq\_tdr} + s^* \cdot 40(ft/sample).$$

Step 6: Predict the US and DS data rates.

Let the US rate-length function for 26AWG loops be $Rate_{US}(L)$, and the DS rate-length function be $Rate_{DS}(L)$, then the US rate, denoted as *US_Rate*, and the DS rate, denoted as *DS_Rate*, are

$$US\_Rate = Rate_{US}\left(L_{eq\_us}\right)\text{, and}$$

$$DS\_Rate = Rate_{DS}\left(L_{eq\_ds}\right).$$

V. Results

The present invention has been tested on several different CO modems that have TDR functionality. The results from one of the modems are given below. All of the loops tested are listed in Table 2. The total number of loops is 1291. All loops are straight loops with a single section or up to four sections. Both American wires and European wires are tested. Fig. 12 shows the estimation results on equivalent-TDR length, with (a) showing the estimated length versus theoretical length, which is calculated using Table 1, and (b) showing the distribution of the estimation error. Fig. 13 shows the estimation results on US data rate, with (a) showing the estimated US rate versus the measured data rate, i.e. the actual data rate when connecting a CO-CPE modem pair using the given loop, and (b) showing the distribution of the error on US rate estimation. Fig. 14 shows the estimation results on DS data rate, with (a) showing the estimated DS rate versus the measured data rate, i.e. the actual data rate when connecting a CO-CPE modem pair using the given loop, and (b) showing the distribution of the error on DS rate estimation.

Table 2: A list of tested loops

No. of sections	Configuration	Length	No. of loops
1-section	26AWG (L)	L = 1kft - 22kft / 1kft	22
	26AWG (L)	L = 1kft - 18kft / 100ft	171
	0.4 mm (L)	L = 200m - 3500m / 50m	67
	0.5 mm (L)	L = 200m - 5000m / 50m	97
2-section	26AWG (L1) + 24AWG(L2)	L1, L2 = 1kft - 9kft / 500ft	289
	0.4mm(L1) + 0.5mm(L2)	L1, L2 = 200m - 3000m / 200m	225
	ETSI #3 (L=length of 0.4mm section)	L = 50m - 3000m / 50 m	60
3-section	26awg(L1)+24awg(L2)+26awg(L3)	L1, L2, L3 = 1kft - 9kft / 2kft L1+L2+L3<=18kft	90
	ETSI #6 (L=length of 0.4mm section)	L = 50m - 3000m / 50 m	60
	ETSI #7 (L=length of 0.4mm section)	L = 50m - 3000m / 50 m	60
4-section	26awg+24awg+26awg+24awg	The length of each section is randomly chosen. The total physical length is less than 18kft.	30
	ETSI #4 (L=length of 0.4mm section)	L = 50m - 3000m / 50 m	60
	ETSI #5 (L=length of 0.4mm section)	L = 50m - 3000m / 50 m	60

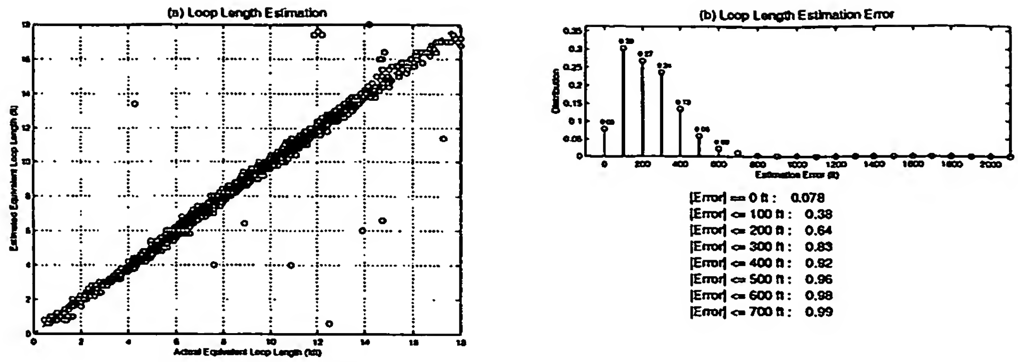

Figure 12: Loop length estimation results. (a) Comparison between the actual and the estimated lengths. (b) Distribution of the estimation error.

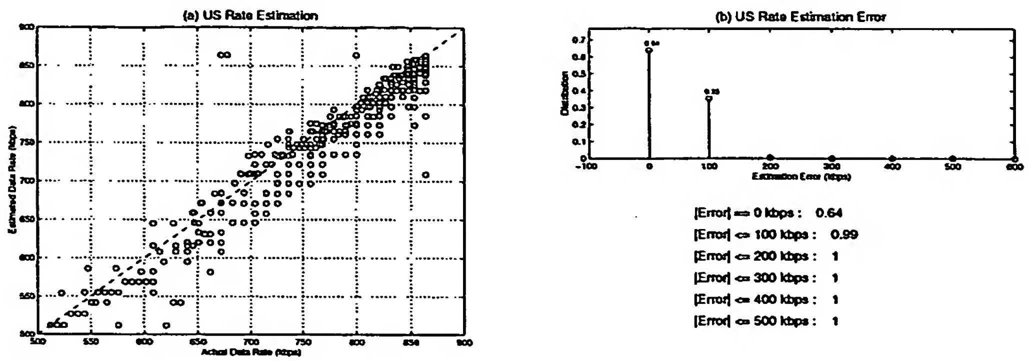

Figure 13: US data rate estimation results. (a) Comparison between the actual and the estimated US date rates. (b) Distribution of the estimation error.

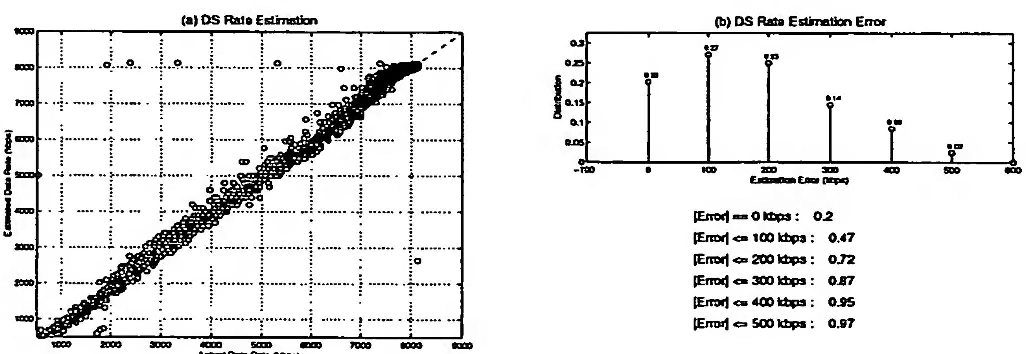

Figure 14: DS data rate estimation results. (a) Comparison between the actual and the estimated DS date rates. (b) Distribution of the estimation error.

Fig. 12, 13 and 14 indicate (1) the equivalent-TDR length estimation error is less than 500ft for 96% of the loops; (2) the US data rate estimation error is less than 100kbps for 99% of the loops, and (3) DS data rate estimation error is less than 500kbps for 97% of the loops.

VI. Conclusions

The present invention relates to predicting the obtainable data rate from a given subscriber line for a potential service, such as but not limited to DSL service, based on the measured TDR echo without explicitly estimating the topology of the line.

Accordingly, aspects of the invention relate to:

A method for estimating the obtainable data rate of a given subscriber line from the TDR echo via estimating the equivalent 26AWG straight loop. Broadly, the gauge of the equivalent loop is not necessary to be 26AWG, it could be any gauge. The only requirement is the data rate of the selected gauge should be known.

A method for evaluating whether two loops with different gauges have the same TDR echo (shape and amplitude) via compensating the time delay difference between them. The waveform of a TDR echo, including the shape and amplitude, reflects the attenuation of a subscriber line. Thus this method makes it possible to evaluate whether two loops have the same attenuation from the TDR echo. This method is not only applicable to single gauge straight loop, but also multi-section loops with different gauges.

A method for estimation the physical length of a given loop based on the equivalent-TDR loop length and time shift between the measured echo and the echo from the equivalent loop.

Although the present invention focuses on multi-section straight loops, it has the potential to be applied to loops with bridged taps.

The attached documents provide additional non-limiting implementation examples and algorithmic details that support one or more of the exemplary embodiments of this invention.

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